higher fieldstrengths at the junction, and, consequently, to a larger number of available avalanche nuclei. This large number will prevent large current concentrations to be formed at separate places of the junction and therefore, only normal noise will be found in germanium p-n junctions during breakdown.

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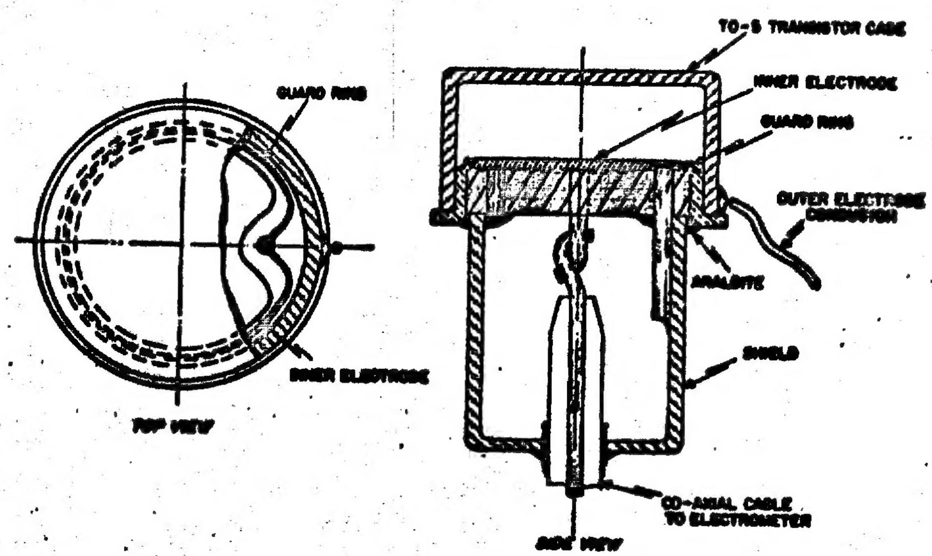

Fig. 1—Ionization chamber.

A Simple Ionization Gauge For Ionizing Radiation Studies on Transistors

It has been observed that one type of radiation damage in semiconductor devices appears to depend upon the ionization produced by radiation within the device encapsulation, rather than upon the bulk damage produced in the semiconductor.[1] In the study of such effects it may be desirable to measure this ionization rate when the device is subjected to ionizing radiation of various types and energies. In the case of certain types of radiation (*e.g.*, monoenergetic γ-radiation in the Mev range), this presents no problem, as attenuation of the beam by the device encapsulation, and wall effects, have only a minor effect on the dose rate delivered inside the transistor can. In the case of β- and X-radiation, whose spectra may be continuous and rather complicated, the effect of the encapsulation may be quite large and difficult to evaluate. This measurement problem can be solved by designing an ionization gauge enclosed by a shield similar in material and thickness to a transistor can. A further requirement is that this radiation-measuring device be no larger than the semiconductor device under study to reduce measurement errors caused by inhomogeneity of the radiation field, such as exists at positions very close to a radiation source. An ionization gauge constructed from a transistor encapsulation, as described below, meets these requirements.

The construction of such a gauge from a transistor TO-3 can is illustrated in Fig. 1. (A similar unit has been constructed from a TO-18 can.) After removing the cover and semiconductor chip, the metal header platform was ground off, and the three leads cut off flush with the exposed glass surface. A 0.2″ diameter disk of 0.005″ steel shim was then centered on this surface and held in place by soldering to one of the leads. A notch on one side exposes one of the remaining leads. A circular guard ring of 0.005″ gold wire was placed on the glass surface between the disk and the circumference of the header, and held in place by araldice which was allowed to flow under it. This guard ring was connected electrically to the exposed lead. The disk was also reinforced in position by allowing araldite to run under its edge. The cover, which acts as the second plate of the ionization chamber, was then replaced by soldering to the header in a dry nitrogen atmosphere.

The external circuit consists simply of a series battery and electrometer (*e.g.*, Keithley 600A) connected between the ionization chamber plates. The guard ring is connected to the battery-electrometer interconnecting lead, and this point is grounded. It is convenient to use a coaxial cable for connection to the guard ring and inner electrode. A tubular piece of copper shim was cemented to the external surface of the glass header and connected to the outer conductor of the coaxial cable as illustrated in Fig. 1. This serves the dual purpose of intercepting leakage currents flowing across the bottom surface of the header, and of eliminating the air path between the two ionization chamber electrodes external to the chamber itself. A coaxial connector at the electrometer input then completes the shielding of the inner electrode from the high-potential lead connecting the battery to the transistor cover. The latter is run as a separate wire alongside the coaxial cable. Care must be taken in the selection of the coaxial cable as small cable movements generate significant voltages between the outer and inner conductors which render the measurement of small dc currents difficult. Of a selection of cables, RG 62A/u was found to exhibit the effect least.

Calibration may be carried out by means of a cobalt-60 gamma source of known strength, as the attenuation of the radiation by the transistor can is only about 1 per cent. Battery polarity is unimportant since the chamber is effectively of parallel-plate geometry. Thirty volts produced current saturation at dose rates up to about 10,000 rads/hr, with the voltage required for saturation increasing as the square root of dose rate. Linearity of ionization current with dose rate appears to hold very well up to at least 100,000 rads/hr, which was the maximum obtainable from our source. Sensitivity was approximately 10^{-14} A/rad/hr. In the absence of radiation, leakage current is of the order of 10^{-14} A which is completely negligible except when measuring very small dose rates. At a fixed dose rate, ionization current was measured as a function of gauge temperature, and found to be almost independent of this parameter in a range of 20°C above and below room temperature.

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L.O. disturbs
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[1] Peck, et. al., "Surface effects of radiation on transistors," *Bell System Tech. J.*, vol. 42, p. 95, January, 1963.

Motion Sensing by Optical Heterodyne Doppler Detection from Diffuse Surfaces

Optical Doppler sensing using the highly coherent gas laser in an optical heterodyne system has been accomplished using moving (specular) mirrors over significant path lengths where most of the engineering problems involved are by now fairly well understood.

However, the use of diffuse surfaces as the targets impose additional requirements and restrictions on the optics of the system. When the coherent laser beam is reflected from a diffuse surface, there is an effective reduction in its spatial coherence if the received signal pattern is resolved at the detector—*i.e.*, when the receiver is not viewing the return signal as a point source. Without properly restricting both the receiver aperture and the transmitted beamwidth, the

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Doppler return signal from the target can be lost completely by random phase cancellations when mixed with the local oscillator contribution.

When the coherent laser beam is incident on a diffuse surface the light is reflected back by many random scattering elements typically separated by at least an optical wavelength (10^{-6} m). However, since the incident light possesses a high degree of both spatial and temporal coherence, the particles of the target will not only reflect the light in accordance with Lambert's law, but will also behave in the manner of a random phased array to produce a diffraction pattern. This gives the familiar "granular" effect which constitutes a standing diffraction pattern of many needle-like lobes projecting out from the reflecting surface.[1,2]

To preserve the phase coherence in a Doppler shifted return resulting from target motion it becomes necessary to reduce the receiver aperture so that only a few of these diffraction lobes are incident at the detector's photosurface. When the receiver is operating at its diffraction limit a single such element will obtain, for

$$\theta_r = \frac{1.22\lambda}{D_r} \tag{1}$$

where θ_r = linear angular field of view of the receiver, D_r = Receiver aperature diameter and λ = the laser wavelength.

However, for the case of an unrestricted laser beam spot diameter at the target, the corresponding receiver aperture may become so small as to make it impossible to detect the received signal.

But by making the angle that the target spot size subtends at the distance R from the receiver equal to the diffraction limited angle θ_r, the following criterion may be applied:

$$\frac{1.22\lambda}{D_r} = \theta_r = \frac{S_t}{R} \tag{2}$$

where S_t = the beam spot diameter at the target. It is apparent then that D_r can be increased as S_t is reduced. This can be achieved by placing a lens after the laser that is large enough for near field operation at distance R; and this will permit focusing the beam to a small spot at the target. For example, if D_r = 10 cm is desired for a range of 1 mile, the required spot size at the target is 1.24 cm. This can be done by using a transmitting lens of the same size as the receiver aperture D_R. The system shown in Fig. 1 applies this principle where reciprocal transmitter-receiver optics are used for simplicity and ease of alignment.

Using a He:Ne laser operating at 6238 Å a power output of 1.3 mw in the TEM_{00} mode was obtained. Accounting for beam splitter and lens losses the actual transmitted power was 0.6 mw.

With this power it was possible to obtain useful Doppler signal at distances up to 180 feet with a Scotchlite semicooperative diffuse target; and 35 ft with white bond paper.

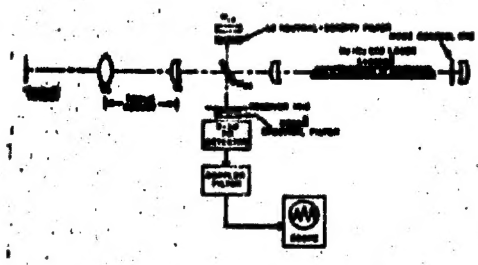

Fig. 1—Optical heterodyne system for Doppler detection from a diffuse target.

TABLE I

Range	Target	S/N from Doppler Filter	Average Doppler Freq.	Filter B.W.
10 ft	White bond paper	20 db	20 KC	5 KC
35	White bond paper	12	20 KC	5 KC
58	Scotchlite	27	20 KC	5 KC
180*	Scotchlite	17	20 KC	5 KC

* For this range a folding mirror was placed at the far end of the dark tunnel.

The maximum aperture used for D_R was 3 cm; and the focused beam size at 35 ft was a fraction of a millimeter.

Table I is a tabulation of the results obtained in a 113 ft dark tunnel.

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[1] J. D. Rigden and E. I. Gordon, "The granularity of Scattered Optical Maser Light," Proc. IRE, vol 50, p. 2367, Nov. 1962.
[2] W. H. Huntley, "New Coherent Light Diffraction Techniques," IEEE Spectrum vol. 1, pp. 114–120, Jan. 1964.

More About Negative-Resistance Circuits

My attention has been called to an error inadvertently made in preparing the drawings for Figs. 1, 2, and 3 of my discussion of Kallmann's negative resistance circuit[1,2] The upper end of the resistance R_c in the collector circuit of each transistor should connect to the emitter of the other transistor, instead of to the emitter of the same transistor.

Since the corrected circuits and the erroneous circuits are similar versions of the Eccles-Jordan circuit, it is of interest to compare the performance of the circuit of Fig. 2 (September) with that of the corrected circuit, which is the exact equivalent of the Kallmann circuit. In the use of the Kallmann circuit in the production of a current-stable negative resistance between the two emitters, the circuit can be represented by the common-emitter equivalent circuit shown herewith in Fig. 1. This circuit is assumed to be completely symmetrical, and R_B' represents the collector-base coupling resistance, made up of the parallel combination of R_B of Fig. 2 (September) and the resistance of the forward-biased coupling diode. The base currents are expressed in terms of the input current I by the relation $I_{b1} = -I_{b2} = IR_C/(R_C+R_B'+h_{ie})$. The input resistance V/I can be determined by inspection of Fig. 1 if it is noted that the voltage V across the input terminals is the sum of the voltages across $1/h_{oe}$ and R_C in either branch of the circuit. If negligible terms are dropped,

$$R_i = \frac{V}{I} \approx -\frac{h_{fe}}{2h_{oe}} \cdot \frac{kR_C}{kR_C + h_{ie}} \tag{1}$$

in which $k = h_{ie}/(h_{ie}+R_B')$. The equivalent circuit for the circuit of Fig. 2 (September) is of the same form as Fig. 1, but R_C appears in parallel with $1/h_{oe}$, instead of in parallel with h_{ie} and R_B', and $I_{b1} = -I_{b2} = I/2$. The input resistance is

$$R_i = \frac{V}{I} \approx -\frac{h_{fe}}{2h_{oe}} \cdot \frac{R_C}{R_C + 1/h_{oe}}. \tag{2}$$

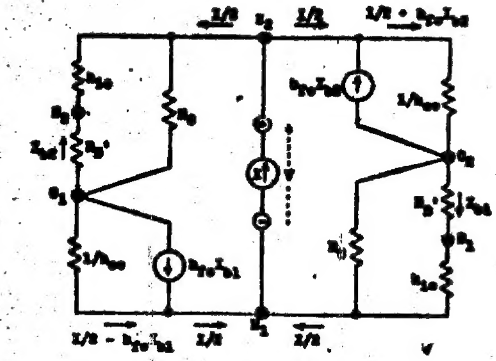

Fig. 1.

Because k/h_{oe} is usually considerably larger than h_{ie}, (1) and (2) indicate that the Kallmann circuit is superior to the pseudo-Kallmann circuit in that the current-stable negative resistance which it produces is greater in magnitude. Furthermore, because a smaller fraction of the input current I flows in each base in the Kallmann circuit, collector-current cutoff occurs at a higher value of I. The Kallmann circuit therefore has the additional advantage that the negative-slope portion of the current-voltage characteristic extends over a greater range of current and voltage. Eq. (1) shows, however, that it is important, in the Kallmann circuit, to minimize the coupling resistance R_B', which reduces the value of k below its maximum value of 1.

When the collector-to-collector port of transistor Eccles-Jordan circuits is used to obtain a voltage-stable negative resistance, one or the other collector junction becomes forward-biased at some value of impressed collector-to-collector voltage that exceeds the maximum magnitude of impressed voltage at which the resistance is negative. To prevent excessive collector current when this occurs, it may be desirable to use, in series with each collector, a diode that conducts only forward collector current.

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[1] Proc. IEEE, (Correspondence) vol. 52, pp. 1068–1069; September, 1964.
[2] H. E. Kallmann, "A simple 1-AC negative-impedance converter," Proc. IEEE, vol. 52, pp. 199–200; February, 1964.